HEWLETT PACKARD

APPLICATION BULLETIN 78

Low-Cost Fiber-Optic Links for Digital Applications up to 150 MBd

(I) THE HFBR-2406/16 HIGH-PERFORMANCE COMPONENT

The HFBR-2406 and HFBR-2416 are high-speed, low-cost linear light-to-voltage converters with typical bandwidths of 125 MHz. These components can be used to make fiber-optic links for both analog and digital applications. Since the range of possible uses are so varied, this Application Bulletin concentrates on a specific digital application. The application is one of the most prevalent for the HFBR-24X6: the transmission of encoded digital signals, otherwise known as run-length limited* data.

The HFBR-0400 component family's inexpensive, one-piece plastic package allows engineers to construct low-cost high-performance fiber-optic links. All devices in the HFBR-0400 family, including the HFBR-24X6, are available with optical ports that are compatible with the industry standard SMA and ST** fiber-optic connectors. Components that are compatible with the SMA connector are denoted by a "zero" in the third digit of their part numbers. If the ST connector is to be used, the component part number should contain a "one" in the third digit. For example, the equivalent of the high performance HFBR-2406 SMA-compatible receiver with the ST connector option is the HFBR-2416.

The addition of the HFBR-24X6 receiver to the low-cost 0400 component family opens new avenues for designers. They can now develop fiber-optic links that meet tough cost and performance objectives. The wide bandwidth of the HFBR-24X6 allows high-speed fiber-optic links to be built at lower prices than was formerly possible. Engineers can exploit the high performance of the HFBR-24X6 in other ways as well. For instance, the wide bandwidth of the linear light-to-voltage converter can be reduced by a low-pass filter to improve the sensitivity of the fiber optic receiver in lower-speed applications. The HFBR-24X6 accommodates a larger optical signal than other HFBR-0400 fiber-optic receivers before it begins to overload. This improvement in the overload characteristics of the 24X6 was achieved with no significant reduction in the ultimate sensitivity when compared to the existing HFBR-24X4 receiver. The increased optical input power tolerated by the HFBR-24X6 allows it to function at short fiber lengths with large values of launched optical power. When the receiver can tolerate higher optical power, a longer cable is possible before attenuation reduces the light to the sensitivity limit of the receiver. The increased dynamic range of the HFBR-24X6 will thus permit greater optical link length for any given fiber attenuation.

*Run-length limited means a limit on the number of consecutive symbols in the same state.

**ST is a trademark of AT&T Technologies.

(II) APPLICATIONS FOR 820 nm LED BASED FIBER OPTIC LINKS

The 820 nm LED technology used in the HFBR-0400 family of components can be used in conjunction with the HFBR-24X6 receiver to construct digital fiber-optic links that transmit data at speeds up to 150 MBd. The length of the fiber light guide that can be used with the HFBR-24X6 is restricted by the receiver sensitivity at low data rates. As the data rate is increased a phenomenon known as chromatic dispersion begins to limit the maximum distance. This distance restriction results from the interaction of the 60 nm-wide spectrum emitted by the LED and the various velocities of light, at wavelengths near 820 nm, in silica. Since the velocities of light at various wavelengths near 820 nm are different, the optical pulses sent by the LED are dispersed or spread out in time as they travel down the light guide. A chromatic dispersion null exists at a wavelength of 1300 nm in silica glass. If an LED were operated at the chromatic dispersion null the pulses would experience the minimum broadening as they traveled through the fiber. This is due to the nearly equal propagation velocity for all the wavelengths transmitted through the silica light guide by the long-wavelength emitter. Figure 1 illustrates the effect of the LED center wavelength and spectral width on the chromatic dispersion. An 820 nm LED with a 60 nm emission spectrum is shown to produce a larger change in the arrival time of the light pulses than a 1300 nm LED with an 100 nm spectral width.

When selecting a fiber the designer should be aware of how the bandwidth-length product, expressed in MHz*km, was determined. The bandwidth of a fiber measured using a narrow spectrum emitter, such as a laser diode, is related

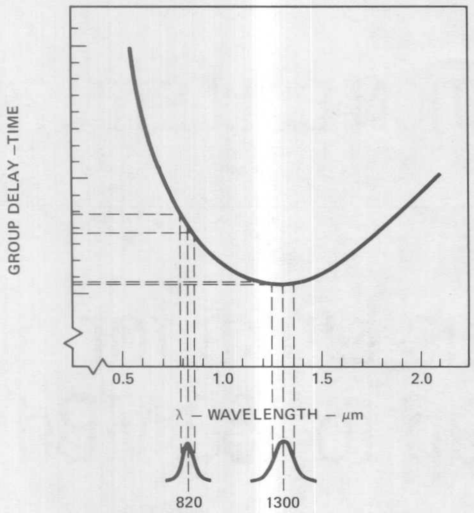


Figure 1.

$$\text{B.W.} = \left[\frac{1}{\left(\frac{1}{\text{B.W. Modal}}\right)^2 + \left(\frac{1}{\text{B.W. Chromatic}}\right)^2} \right]^{1/2} \quad (1)$$

to the various possible modes of light propagation that can exist in a fiber. This is referred to as the fiber's modal bandwidth. The modal bandwidth will be greater than the chromatic bandwidth which dominates when an LED is used. To determine the overall optical bandwidth of a fiber, the modal and chromatic bandwidths must be combined as an rms sum as shown in Equation 1. In LED-based systems the wavelength, spectral width and response time of the emitter used as the fiber-optic transmitter will affect the final system bandwidth. Thus, to understand how a fiber will work with an LED, the type of source used to measure the manufacturer's stated bandwidth must be known. HP HFBR-AWSXXX 100/140 μm fiber-optic cable has a typical optical bandwidth-length product of 40 MHz·km. This value represents the performance of the HP fiber with an 820 nm LED emitter that has a 60 nm spectral width. The 40 MHz·km typical bandwidth length product of HP fiber results from the combination of the modal and chromatic bandwidths.

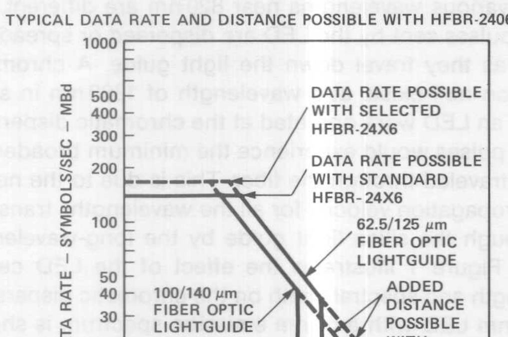


Figure 2.

The typical distances and data rates possible with 820 nm LED emitters and the HFBR-2406/2416 receiver are shown in Figure 2. Note that the data rate versus distance for 100/140 μm and 62.5/125 μm graded-index fibers are both shown in the figure. [1] [2] [3] [4]

If greater distances or higher speeds are required, other options such as 1300 nm LEDs or laser diodes can meet these objectives. If the system requirements fall to the left of the curves shown in Figure 2 the design goals can be achieved using an 820 nm LED and the HFBR-24X6 for a substantially lower cost than possible with these other technologies. The inexpensive 820 nm LED technology offers the designer a cost-effective solution sufficient for many short-distance applications at data rates in excess of 100 megabaud.

APPLICATIONS FOR 820 nm LED BASED SYSTEMS USING HFBR-2406/16 INCLUDE:

— CPU to disc interface.
— CPU to monitor interface.
— CPU to peripheral interface.
— Optical data bus applications.
— Graphics workstation to host computer interface.
— Wide dynamic range, long distance, medium speed LAN applications.
— High speed point to point data links.

(III) ADVANTAGES OF RUN LIMITED CODE

The data is coded to prevent the digital information from remaining in one of the two possible logic states for an indefinite period of time. The coded data allows the fiber-optic receiver to be ac coupled. Without encoding, the fiber-optic receiver would have to detect dc levels to determine the proper logic state during long periods of inactivity. This situation occurs when there is no change in the transmitted data. Ac-coupled fiber-optic receivers tend to be lower in cost, are much easier to design, and contain fewer components than their dc-coupled counterparts.

Direct coupling decreases the sensitivity of a fiber-optic receiver since it allows the low-frequency flicker noise from the transistor amplifiers to be presented to the comparator input. Any undesired signals coupled to the comparator will reduce the signal-to-noise ratio at this critical point in the circuit, and reduce sensitivity of the fiber-optic receiver.

Another problem associated with direct-coupled receivers is minimizing the accumulation of dc offset. With direct coupling, the gain stages multiply the effects of undesirable amplifier offsets and voltage drifts due to temperature changes, and apply them to the comparator. Increases in the dc offset applied to the comparator result in reduced sensitivity of the fiber-optic receiver. The dc offset at the comparator can be referred to the optical input of the receiver by dividing by the receiver gain. This division refers the dc offset at the comparator to the receiver input where it appears as a change in optical power that must be exceeded before the receiver will switch states.

Another advantage of run-limited coding is related to timing recovery. If NRZ data were transmitted over a serial fiber-optic link the data could be in the logic "1" or "0" state for an indefinite period of time. When NRZ data remains in a particular state no transitions occur and the fundamental frequency of the data is dc. This lack of power at the fundamental frequency of the data eliminates the reference signal needed by the timing recovery circuits required to

clock the received information. If an optical link is to transmit NRZ data, a clock signal must be sent on a separate fiber-optic link to synchronously detect the incoming serial data.

The particular run-length-limited code chosen must be considered carefully since it will affect the bandwidth required by the serial communication channel. A complete discussion of all run-limited codes is beyond the scope of this publication. If you desire additional information regarding various coding schemes, you will find it in the Hewlett-Packard Optocoupler and Fiber Optic Applications Handbook. Without becoming too involved in the complexity of encoding selection, a quick comparison will now be made between two commonly recognized approaches to this problem.

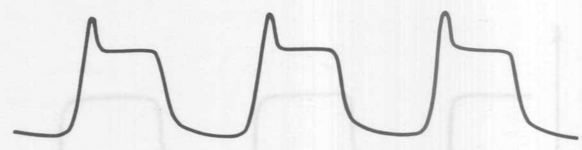
One of the most familiar run-limited codes is Manchester. Manchester is very popular since it can be encoded and decoded with relatively simple circuits. Manchester works well in ac-coupled systems since it has a 50% duty factor and two pulses or symbols for each bit transmitted. This simplifies the design and implementation of the timing recovery function since Manchester code has only two consecutive symbols without a transition, or a run limit of two. A drawback of Manchester is that two symbols must be sent for each data bit encoded, thus doubling the fundamental frequency that must pass through the information channel. Substitution codes have recently been made available in VLSI integrated circuits that function as a general purpose interface between the parallel architecture found in computer-based systems and the serial format required by fiber-optic communication links. The two different substitution codes available in the Am 7968 parallel-to-serial encoder are 4B5B and 5B6B. These two codes have an efficiency of 4/5 and 5/6 respectively which compares to an efficiency of 1/2 for Manchester code. The significance of coding efficiency can be illustrated by an example. If an application calls for the transmission of 100 M bits/second, Manchester code requires that the information channel must pass 200 M symbols/second or 200 MBd. If the more efficient 4B5B code were used, 100 M bits/second could be sent at a speed of (5/4)(100 M bit/sec) = 125 MBd. Similarly, use of 5B6B would allow transmission of this data at a speed of (6/5)(100 M bit/sec) = 120 MBd.

Regardless of the particular coding scheme used there will always be two symbols per cycle. This is true because each half cycle of the maximum fundamental frequency of the data that the communications channel must pass could be either a one or a zero in a binary transmission system.

(IV) DESIGNING WITH FIBER OPTIC COMPONENTS

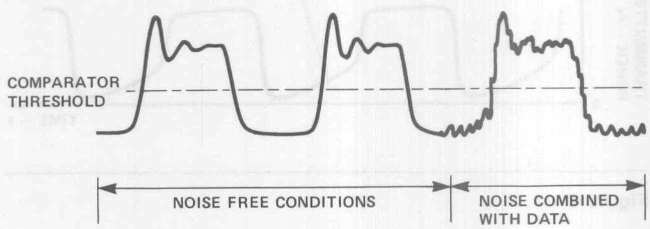
Transmitter Design

Now that the basic issues related to fiber-optic link design have been covered, some specifics related to the design of the optical transmitters and receivers will be discussed in greater detail. To achieve the wide bandwidth performance potential of the fiber-optic medium requires a fast LED and current modulator. The transmitter's pulse-width distortion and optical rise and fall times can be heavily influenced by the driver selected. Readily available off-the-shelf integrated circuit current drivers can be configured with the HFBR-14XX 820 nm LEDs to build high-performance fiber-optic transmitters with a typical pulse width distortion of 1.2 nsec.



OPTICAL OVERSHOOT DUE TO EXCESSIVE PEAKING OF THE LED DRIVE CURRENT

Figure 3a.



RESPONSE OF OPTICAL RECEIVER TO AN EXCESSIVELY PEAKED LED TRANSMITTER

Figure 3b.

To obtain the best performance from any LED and driver combination, two simple techniques known as prebias and drive current peaking should be employed. Prebias, as its name implies, is a small forward current applied to the LED in the "off" or "low" light state. The prebias current prevents the junction and parasitic capacitances from discharging completely when the LED is in the "off" state, thus reducing the amount of charge that the driver must transfer to turn the emitter back on. Peaking is a momentary increase in LED forward current that is provided by the driver during the rising and falling edges of the current pulses that are used to modulate the emitter. If the time constant of the peaking circuit is approximately equal to the minority carrier lifetime of the emitter, the momentary increase in LED current will transfer charge at a rate that improves the rise or fall time of the light output without causing excessive overshoot of the optical pulses. Problems that can result when excessive peaking is applied to the LED are illustrated in Figure 3. The narrow optical overshoot due to excessive peaking of the transmitter causes a narrow electrical output pulse from the fiber-optic receiver that must now be damped. Even if the receiver amplifiers were critically damped the electrical undershoot resulting from excessive peaking of the emitter can reduce the sensitivity of the fiber-optic link. This electrical undershoot can combine with noise from the amplifiers so that the sum of these two voltages exceeds the decision threshold of the comparator, which converts the low-level analog output of the fiber-optic receiver back to logic-compatible digital signals. Excessive peaking during the turn-off of the emitter can cause additional problems. Too much reverse current during the turn-off transition will reverse-bias the LED, seriously degrading the turn-on time.

Because the current driver is attempting to modulate a device that has a nonlinear characteristic, a driver with a nonuniform source impedance can actually provide some useful advantages. LED emitters are characteristically harder to turn off than they are to turn on. This difficulty manifests itself as a phenomenon commonly referred to as the long-

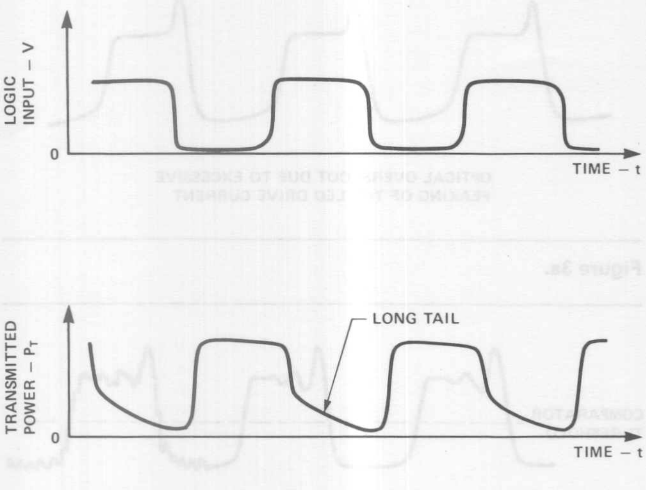


Figure 4.

tailed response. An example of the long-tailed response is shown in Figure 4. The long-tailed response is most evident when a simple series switch is used to control the current through the LED as shown in Figure 5. A shunt drive configuration, which turns the LED off when the driver transistor saturates, significantly improves the performance of the LED transmitter. Shunt drive reduces pulse-width distortion and the magnitude of the slow tail by providing a lower impedance path for charge stored in the LED junction. Without this low-impedance path the emitter would turn off slowly since the LED would continue to produce light until the diode junction is discharged.

Readily available high current line driver ICs can be used to current-modulate the LED in TTL interface applications. Ordinary TTL gates generally do not have sufficient sink and source current capability to directly drive the LED emitter. A simple high speed transmitter can be constructed by connecting the totem pole output of a TTL line driver to the LED as shown in Figure 6. In this configuration the pull-down transistor turns the emitter off and the pull-up transistor turns the device on. The lower impedance and higher current handling capability of the saturated pull-down transistor can be used as a very effective method of transferring charge as the LED dynamic impedance increases during turn-off. The somewhat higher output impedance of the pull-up half of the driver ensures that the LED isn't over peaked during the less difficult turn-on transition. This

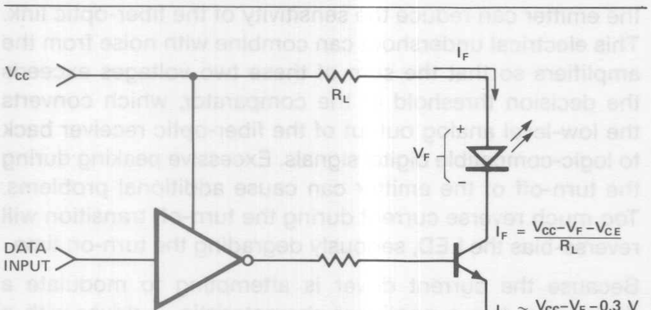


Figure 5.

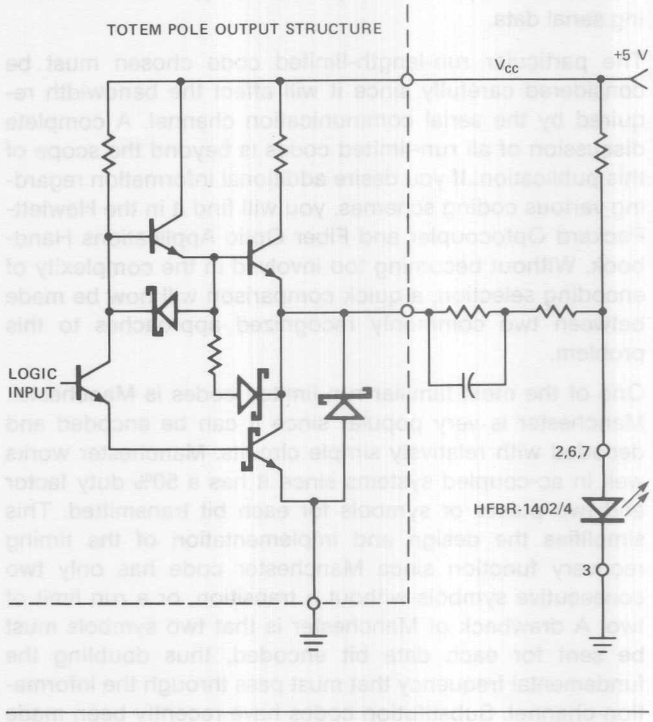


Figure 6.

asymmetric current handling capability combined with an LED geometry that takes advantage of such nonuniform peaking can substantially reduce the pulse-width distortion that results from the long-tailed response. When the TTL-compatible line driver and LED are configured as shown by the schematic in Figure 6 the improvement in the optical output waveform is as shown in Figure 7. Care should be taken when selecting a TTL line driver since this device must have rapid rising and falling edges, and be able to handle the relatively large source and sink currents required to drive an LED. In addition to sufficient current handling capacity, the driver I.C. should have minimal difference between t_{PHL} and t_{PLH} as the spread between these gate propagation delays will adversely affect the pulse-width distortion present in the optical output of the transmitter.

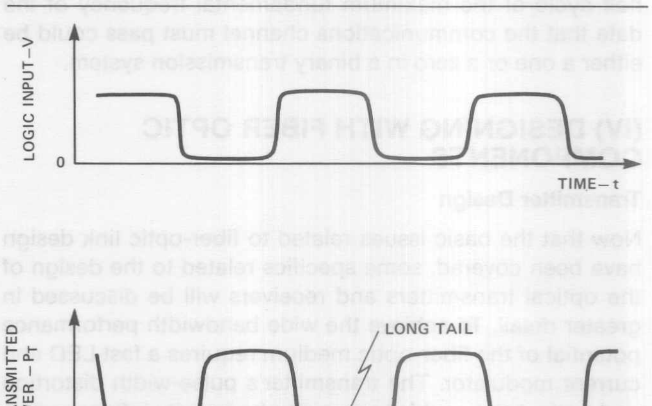
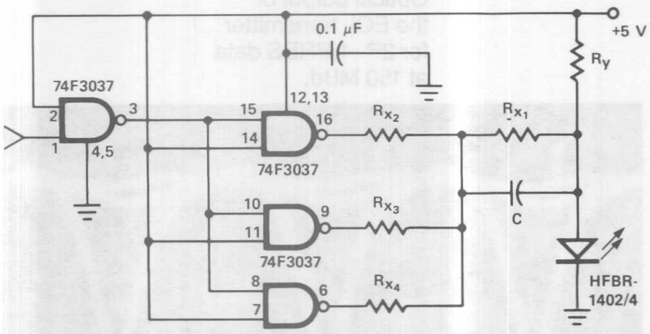


Figure 7.



THIS DRIVER CRT WILL WORK TO MAX I_{FDC} = 60 mA OF THE LED EMITTER.

PEAK OUTPUT OPTICAL POWER MEASURED OUT OF 1 M OF CABLE

		I_F = 60 mA	T_A = 25° C
Fiber Cable	NA	HFBR-1402	HFBR-1404
HP 100/140 µm	0.3	-11.5 dBm	-7.1 dBm
62.5/125 µm	0.28	-16.5 dBm	-12.2 dBm
50/125 µm	0.20	-21.9 dBm	-17.5 dBm

Figure 8.

$$R_y = \frac{(V_{CC}-V_F) + 2.84\,(V_{CC}-V_F-1.6\ V)}{I_{FON}} \qquad (2)$$

$$R_x = \left(\frac{R_y}{2.84} - 10\ \Omega\right)$$

$$R_{x_1} = \frac{R_x + 10\Omega}{2}$$

$$R_{EQ_2} = R_{x_1} - 10\Omega$$

$$C = \frac{2.0\ n\ SEC}{R_{x_1}}$$

$$R_{EQ_2} = R_{x_2} \parallel R_{x_3} \parallel R_{x_4}$$

Where

$$R_{x_2} = R_{x_3} = R_{x_4} = 3\,R_{EQ_2}$$

The transmitter shown in Figure 8 is compatible with TTL logic and is suited for data with a maximum fundamental frequency of 60 MHz, which implies a symbol rate of 120 MBd. The design rules for the LED driver shown in Figure 8 are shown in Equation 2. This simple TTL-compatible fiber-optic transmitter has a typical rise/fall time of 3 ns. Additional measurements were taken for temperatures between -40 to +85° C. They indicate that using components with 5% tolerance in the network between the 74F3037 and the HFBR-1402/4 LED resulted in a typical pulse-width distortion in the optical output of 15% for a data rate of 120 MBd.

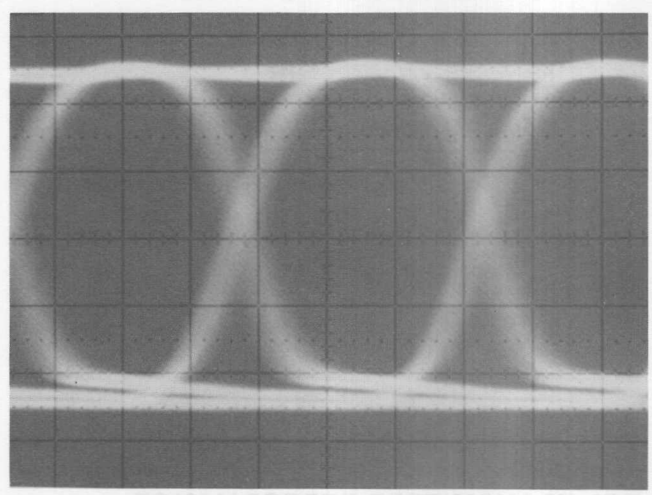
TESTING FIBER OPTIC SYSTEMS

Pseudo-random-bit-sequence (PRBS) generators are very useful for testing the performance of fiber-optic systems. The pseudo-random data pattern contains long periods of inactivity related to the length of the shift register used to build the PRBS generator. A PRBS generator made up of a 23-bit-long shift register could at any given clock time contain one of 8,388,610 possible data patterns. The number of data patterns possible can be calculated as (2^{23})-1 since the state where all shift register stages contain logic zeros is not allowed. These long periods of inactivity in the data pattern produced by the PRBS generator allow time for parasitic capacitances in the transmitter and receiver to charge. The time required to charge and discharge undesired capacitances in the transmitter and receiver result in pulse-width distortion related to the instantaneous duty factor of the data. If an oscilloscope is clock triggered on the PRBS generator it asynchronously samples the data due to the lack of correlation between the PRBS clock and the time base that generates the horizontal sweep of the scope. When triggered on the PRBS generator's clock the scope will display a signal known as the "eye pattern". The "eye pattern" can be very useful since the width and height of the opening between the data edges defines the time period during which the data is in a valid logic state.

TTL TRANSMITTER PERFORMANCE

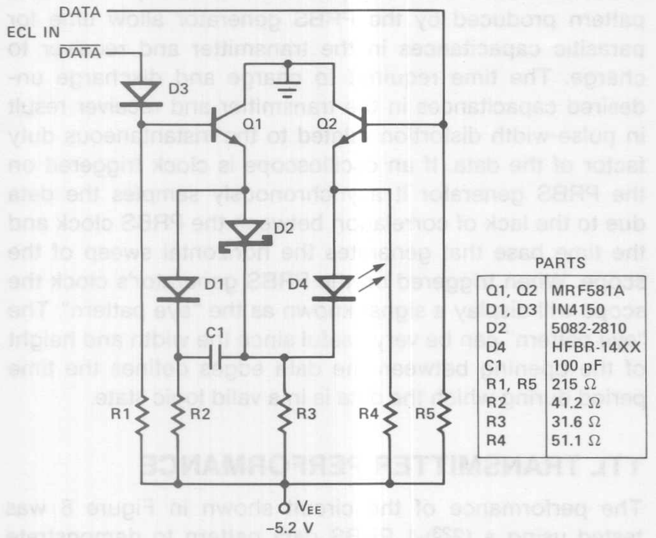
The performance of the circuit shown in Figure 8 was tested using a (2^{23})-1 PRBS data pattern to demonstrate the typical performance of this TTL transmitter. Jitter in the data edges results due to the pulse-width distortion induced by the PRBS data. The "eye pattern" shown in Figure 9 reveals that the HFBR-1402/1404 LED transmitter had a total data-dependent edge jitter of 1.2 ns when driven by the 74F3037 line driver in this 139 MBd PRBS test. This data was taken at an ambient temperature of 25° C and represents the typical performance possible with this simple fiber-optic transmitter. The total pulse-width distortion can be further reduced by using a limited-range potentiometer in place of fixed values of R_y for system applications that are extremely intolerant of symbol-width variations. But for most data communications applications, this transmitter performs adequately at speeds up to 120 MBd using fixed component values.

139-MBd eye pattern for HFBR-1402/1404 LED emitters driven by the 74F3037 TTL line driver.



2 nS/DIV.

Figure 9.



Q1, Q2	MRF581A
D1, D3	IN4150
D2	5082-2810
D4	HFBR-14XX
C1	100 pF
R1, R5	215 Ω
R2	41.2 Ω
R3	31.6 Ω
R4	51.1 Ω

PEAK OUTPUT OPTICAL POWER MEASURED OUT OF 1 M OF CABLE

		I_F = 60 mA	T_A = 25° C
Fiber Cable	NA	HFBR-1402	HFBR-1404
HP 100/140 μm	0.3	-11.5 dBm	-7.1 dBm
62.5/125 μm	0.28	-16.5 dBm	-12.2 dBm
50/125 μm	0.20	-21.9 dBm	-17.5 dBm

Figure 10.

$$R3 = \frac{(V_{OH}-V_{BE}-V_F) - V_{EE}}{I_{F_{DC}}} \quad (3)$$

$$I_{F_{PEAK\ @\ TURN\ ON}} = 2X(I_{F_{DC}})$$

$$I_{F_{PEAK\ @\ TURN\ ON}} = I_{F_{DC}} + \frac{(V_{OH}-V_{D3}-V_{BE}-V_{D1}) - V_{EE}}{R2}$$

$$2(I_{F_{DC}}) = I_{F_{DC}} + \frac{(V_{OH}-V_{D3}-V_{BE}-V_{D1}) - V_{EE}}{R2}$$

$$R2 = \frac{(V_{OH}-V_{D3}-V_{BE}-V_{D1}) - V_{EE}}{I_{F_{DC}}}$$

$$R4 = \frac{(V_{OH}-V_{BE}) - V_{EE}}{I_{F_{PEAK\ @\ TURN\ OFF}}}$$

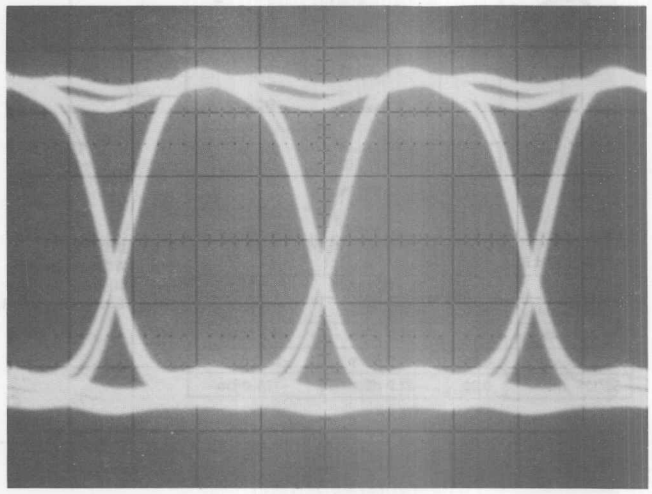
$$C1 = \frac{4\ \text{nsec}}{R2}$$

Where: $V_{OH} = -0.9\,V$; $V_{BE} = 0.9\,V$
$V_{D1} = 0.79\,V$; $V_{D3} = 0.6\,V$

ECL TRANSMITTER PERFORMANCE

If a higher speed ECL-compatible fiber-optic transmitter is needed it can be easily built using the circuit shown in Figure 10. The design rules for this higher-performance fiber-optic transmitter are given in Equation 3. This particular transmitter uses two discrete MRF-581A microwave transistors in conjunction with the HFBR-14X2/X4 LED emitter. It is capable of typical optical rise/fall times of 2.4 nsec. The performance of the ECL transmitter was measured with an HP 81519A 400 MHz optical receiver. Figure 11 shows the optical "eye pattern" that results when a 150 MBd pseudo-random-bit-sequence of (2^{23})-1 is applied to the ECL transmitter.

Optical output of the ECL transmitter for $2^{23}-1$ PRBS data at 150 MBd.



2 nsec/DIV.

Figure 11.

RECEIVER DESIGN

Now that the techniques required to build high-speed fiber-optic transmitters have been explained, emphasis must be placed on the methods necessary for design and construction of the fiber-optic receiver. Figure 12 shows the functional blocks required to interface the HFBR-24X6 light-to-voltage converter to digital logic. The HFBR-24X6 has a low-level analog output related to the incoming optical power by the 7 mW/μW conversion gain of the light-to-voltage transducer. The HFBR-24X6 needs additional external gain stages to increase the amplitude of its output before it can interface to any of the standard logics like TTL or ECL. The output voltage of the HFBR-24X6 is proportional to the received optical flux. Since the received optical power changes as a function of the fixed optical losses and as a function of fiber-optic link length, some provision must be made to accommodate the change in the output voltage of the light-

FIBER OPTIC RECEIVER BLOCK DIAGRAM

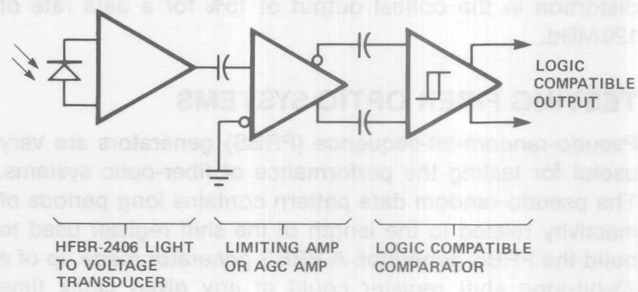


Figure 12.

to-voltage transducer. An amplifier with AGC or a limiting amplifier is needed to accommodate the wide range of output voltages that are possible under various fiber link operating conditions. In the following example, calculations show that the output voltage of the HFBR-24X6 could range from a minimum of 2.9 mVpp to a maximum of 1.74 Vpp. This output voltage range is for worst case conditions at a BER less than or equal to 1×10^{-9} when operating between the -40 to +85°C temperature range of this component.

The peak-to-peak signal-to-rms noise ratio needed at the comparator input for a BER of 1×10^{-9} = 12:1.

This implies an extinction to peak (peak-to-peak) change in the received optical flux of 12●(rms noise) will be required. Thus the peak-to-peak to rms noise ratio required by the fiber-optic receiver for a BER of 1×10^{-9} = 12:1.

The noise floor of the HFBR-24X6 is -42.7 dBm rms typical.

-42.7 dBm + [10 log (12/1)] = -42.7 dBm + 10.8 dB = -31.9 dBm pk. Thus -31.9 dBm pk is the minimum received optical power that will yield a BER better than or equal to 1×10^{-9}.

-31.9 dBm implies [antilog (-31.9/10)](1,000) = 0.644 μW min. received optical power for BER better than or equal to 1×10^{-9}.

This min. power of 0.644 μW implies a change in the receiver input from approx. 0 μW to 0.644 μW or a peak-to-peak change of approx. 0.644 μW pp. The minimum output of the HFBR-24X6 thus becomes (0.644 μW pp)(4.5 mV/μW) = 2.90 mV pp.

The HFBR-24X6 overloads at -8.2 dBm worst cast minimum. Overload is specified as Pr max. on the data sheet. Overload is defined as the received optical power at which the output pulses from the HFBR-24X6 are distorted 2.5 ns due to saturation of the transimpedance amplifier used to convert photo-current to voltage.

-8.2 dBm implies [antilog (-8.2/10)](1,000) = 151 μW. Thus the max. allowed power of 151 μW implies a change in the receiver input from approx. 0 μW to 151 μW voltage or a peak-to-peak change of approx. 151 μW pp. Thus max. received optical power of 151 μW implies a maximum output of (151 μW pp)(11.5 mV/μW) = 1.74 V pp.

ERROR RATE VERSUS SIGNAL TO NOISE RATIO

The bit error rate (BER) possible with a fiber-optic link is a function of the difference between the peak-to-peak signal and the RMS noise voltages present at the comparator input. A linear relationship exists between optical power entering the HFBR-24X6 and the voltage output of the fiber-optic receiver, provided that interstage coupling and post amplifiers do not introduce significant distortion. This linear relationship implies that if a peak-to-peak signal voltage 12 times larger than the RMS noise voltage is needed at the comparator to ensure a BER of 1×10^{-9}, then the same ratio will be required at the receiver input. Thus, the difference between the peak-to-peak optical input of light pulses applied to the HFBR-24X6 and the RMS equivalent noise power referred to the optical input must also be 12 to 1. Some confusion exists because changes in the emitter output from the extinction to maximum power are often referred to as peak excursions of the transmitter launched power. This confusion results since the transmitter output is varying from zero light to a maximum or peak light output. The extinction-to-on excursion in the optical output of an emitter is actually a peak-to-peak change in intensity. Figure 13 is a graph of the ratio between peak-to-peak signal power and the rms equivalent noise power of the receiver resulting from extensive reduction of statistical theory that relates the probability of an error to the signal-to-noise ratio.

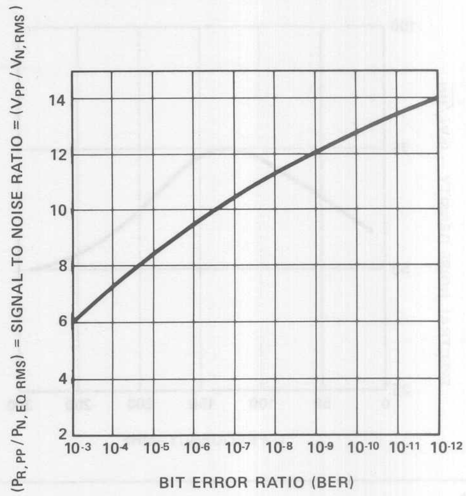


Figure 13.

ADVANTAGES OF HYSTERESIS

The use of hysteresis in the digitizer will not change the signal-to-noise ratio required at the comparator for a particular BER. Hysteresis will, however, introduce a discontinuous response in the receiver that alters the ratio between peak signal level and the RMS noise in stages prior to the comparator. When dual-threshold detection is used the signal-to-noise ratio required at the decision circuit for a particular error rate is unaffected, while the change in the received power level required to switch the state of the comparator is increased in proportion to the amount of the hysteresis. Dual-threshold receivers experience a reduction in sensitivity proportional to the amount of hysteresis used, however, this type of digitizer offers some interesting advantages. Hysteresis can be combined with a differentiator that passes only the high frequencies contained in the data edges. This differentiator and comparator with hysteresis can be combined with a low-pass filter that minimizes the bandpass of the receiver. The advantage of this technique is that all spurious signals other than noise that occurs at the same frequency as the data will be rejected. Hysteresis is used in all the receivers shown in this Applications Bulletin. Use of hysteresis insures that the logic output of the fiber-optic receiver will not toggle in response to the rms output noise voltage of the HFBR-24X6 when no fiber is connected.

LOW PASS FILTERING TO ENHANCE RECEIVER SENSITIVITY

The importance of filtering to eliminate unnecessary receiver bandwidth becomes apparent by studying Figure 14, which shows the relationship between frequency and the spectral noise density of the HFBR-2406/2416. If the fiber-optic link under consideration were intended for operation at 50 MBd (which implies a fundamental data frequency of

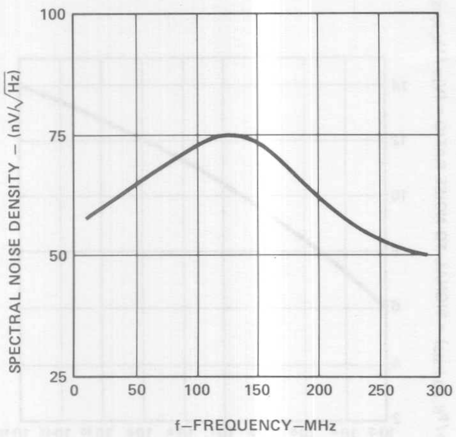


Figure 14.

25 MHz) a substantial increase in receiver sensitivity can be realized. This increase in sensitivity is obtained by filtering out the noise peak that occurs in the HFBR-24X6 at higher frequencies than required for this application.

The selection of the low-pass filter corner frequency should be carefully considered since it is affected by the response of the transmitter, fiber, and receiver. To prevent problems that will cause interference between adjacent pulses of data transmitted over the fiber-optic communications channel, the bandwidth of the entire system from transmitter to receiver must be properly specified. A problem known as intersymbol interference develops when the channel bandwidth is not correctly related to the minimum pulse width of the data that is to be transmitted over the communications system. Insufficient system bandwidth manifests itself as distortion in the receiver output signal at time intervals adjacent to the edges of each symbol. This distortion results in interference between adjacent pulses, which can combine with system noise to create errors. Noise is also directly related to bandwidth. Thus, fiber link performance and BER will degrade if system components are excessively fast. For optimum performance that minimizes the amount of optical power required at the receiver for a given BER, the system bandwidth should ideally be constrained to range between 0.6 to 0.8 times the signaling rate in baud, as shown in Figure 15a. If the bandwidth of the fiber-optic communications channel is excessive, a low-pass filter that restricts the system bandwidth to the amount shown in Figure 15a should be constructed in the fiber-optic receiver at a point ahead of the decision circuit or comparator. For best results the low-pass filter chosen to limit the bandwidth should be a high-order linear-phase type whenever practical. As the frequency increases, the cost and complexity of a linear-phase high-order filter may become excessive. These higher-speed applications will continue to benefit from a simple first-order or second-order RC low-pass filter that will still be practical to implement.

COMPROMISES ASSOCIATED WITH HIGH SPEED 820 nm LINKS

Systems with bandwidths less than (0.6 to 0.8) x (baud) will continue to function since catastrophic failure does not result if these recommendations are violated. Fiber-optic links with bandwidths less than (0.6 to 0.8) x (baud) will have less than ultimate sensitivity but can still be useful at shorter distances. The -32 dBm average sensitivity which is typically possible with the HFBR-24X6 will allow fiber-optic links to operate beyond frequencies within the flat portion of the system's amplitude response. Figure 15b shows an example optical link whose mid-band amplitude response has been normalized to one. If this link were operated under conditions that reduced the total system output to 6 dB below mid-band amplitude, adequate system margin can be shown to exist. This margin, as calculated in Equation 4, is sufficient for low error transmission of 100 MBd data over a 1 km length of 62.5/125 graded index fiber. The HFBR-24X6 receiver has typically demonstrated a BER less than or equal to 10^{-9} at received optical powers of -32 dBm average (-29 dBm peak) at 100 MBd. In this somewhat pessimistic example, where the link sensitivity was assumed to decrease by 6 dB, the ample 2.9 dB margin remaining will ensure that the BER is better than 1×10^{-9}.

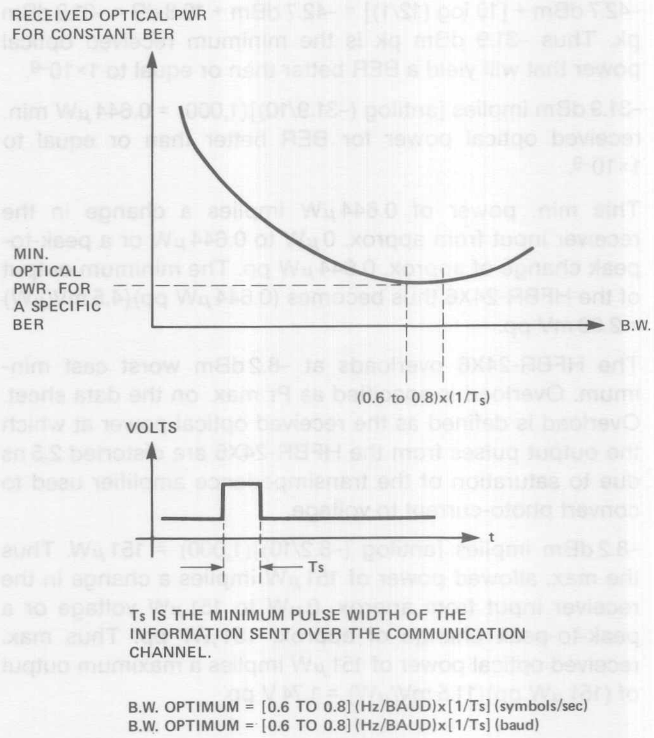


Figure 15a.

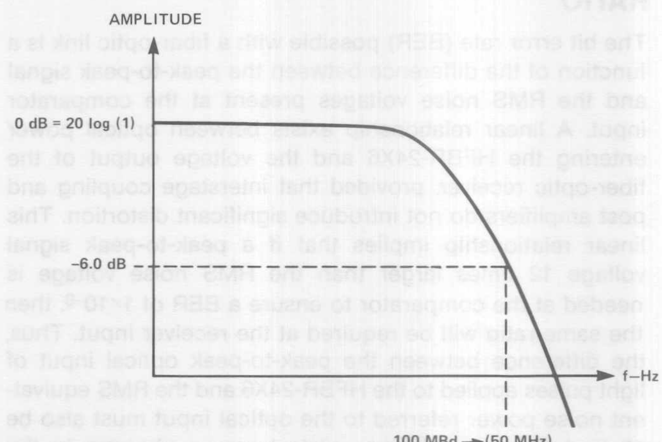


Figure 15b.

OPM = Optical Power Margin (dB) (4)

P_R = Optical Power Required at HFBR-24X6 Receiver for BER $\leq 10^{-9}$. (dBm)

P_T = Transmitter Launched Power. (dBm)

ΔLS = Change in the Link Sensitivity Due to Chromatic Dispersion, and the Response Time of the Transmitter and Receiver. (dB)

$\propto_O \ell$ = Fiber Loss (dB)

$$OPM = -(P_R) + P_T - \propto_O \ell + \Delta LS$$

$$OPM = -(-29) - 16.1 - (4\ dB/km)(1\ km) - 6.0\ dB$$

$$OPM = 2.9\ dB$$

HIGH FREQUENCY CIRCUIT DESIGN

The HFBR-24X6 and each of the amplifiers used in the 10116 are stable gain blocks that have no tendency to oscillate. Although each of these components is individually stable, the combined phase shift and gain that results when they are cascaded might produce oscillation unless proper circuit construction techniques are used. The effect of all the poles that accumulate as the signal is amplified and digitized by the various gain blocks in the receiver results in a very steep high-order roll-off for the overall input-to-output open-loop receiver gain. In essence, the fiber-optic receiver relies on the fact that it is an open-loop system. It has sufficient gain and phase shift to meet the criteria for oscillation if the loop were to be closed. Since closed-loop gain must be kept to less than unity, the attenuation of parasitic and conductive feedback paths must be greater than the open-loop gain of the receiver to prevent undesired oscillations. Parasitic feedback from the high-level logic-compatible output must be kept to a minimum by layouts that physically separate the receiver inputs and outputs. Filtering must be used to ensure that power supply buses do not provide a metallic feedback path that will degrade the stability of the receiver, and a ground plane is recommended to minimize the inductance of supply commons.

When good layout practices are employed, fiber-optic receivers with 150 MBd data rates can be easily constructed using commonly available breadboarding techniques. A sound breadboard technique that can be used for prototyping the HFBR-2406/2416 can be implemented using perforated P.C. boards with holes on tenth-inch centers and a copper-clad ground plane on one side only. Use a small hand-held twist drill holder (pin vise) and a number 32 drill to clear copper away from holes through which the component leads will pass. Do not clear all the copper away between these holes. This copper provides ground connections between each I.C. lead, thus reducing ground-loop size and increasing circuit performance. Install the components on the copper foil side using the component leads for point-to-point wiring interconnections on the insulated side of the board. Production fiber-optic systems can be implemented on ordinary double-sided G-10 printed circuit material or multi-layer boards as long as the layout practices discussed here are observed.

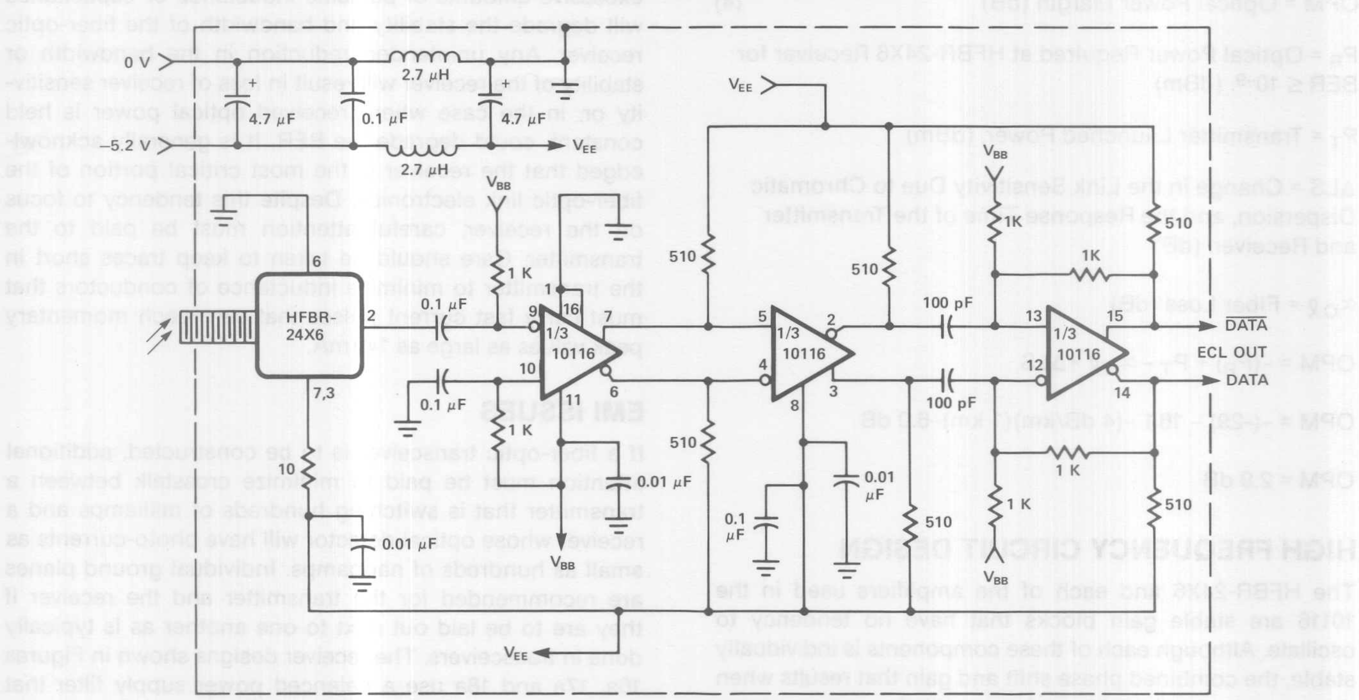
The importance of good construction and layout practices cannot be over-stressed: poor circuit design will seriously degrade system performance. Circuit designs that result in excessive amounts of parasitic inductance or capacitance will degrade the stability and bandwidth of the fiber-optic receiver. Any unintended reduction in the bandwidth or stability of the receiver will result in loss of receiver sensitivity or, in the case where received optical power is held constant, could degrade the BER. It is generally acknowledged that the receiver is the most critical portion of the fiber-optic link electronics. Despite this tendency to focus on the receiver, careful attention must be paid to the transmitter. Care should be taken to keep traces short in the transmitter to minimize inductance of conductors that must carry fast current pulses that can reach momentary peak values as large as 140 mA.

EMI ISSUES

If a fiber-optic transceiver is to be constructed, additional attention must be paid to minimize crosstalk between a transmitter that is switching hundreds of milliamps and a receiver whose optical detector will have photo-currents as small as hundreds of nanoamps. Individual ground planes are recommended for the transmitter and the receiver if they are to be laid out next to one another as is typically done in transceivers. The receiver designs shown in Figures 16a, 17a and 18a use a balanced power supply filter that eliminates noise conducted by both the power and common sides of the voltage source used to power the circuit. This filter should be located between the two power-bus and ground planes of the transceiver to keep transmitter noise out of the receiver. The voltage source used to power the optical transceiver should be connected to the transmitter side of this filter since this half of the circuit is much less sensitive to power supply noise.

Another factor that could degrade the performance of a fiber-optic receiver is environmental noise. The HFBR-2406 combines the PIN diode optical detector and the current-to-voltage converter in a small hybrid package. This minimizes the antenna lengths of the HFBR-2406/2416 at the high-impedance input of the transimpedance amplifier that converts the photo-current to a voltage. This small geometry allows the light-to-voltage converter to have a high electromagnetic interference immunity in excess of 10 V/m. Caution must be exercised, however, to ensure that the metal ferrule of the fiber-optic connector does not act as an EMI source by contacting electrically noisy parts of the system into which it is integrated. If the system in which the fiber-optic link is used is extremely noisy it is recommended that a spring contact be used to ground the metal connector as shown in Figure 19. Another method that reduces the effects of EMI picked up by the fiber-optic connector is to use a connnector with a non-conductive plastic or ceramic ferrule. In some extremely noisy applications the fiber-optic components were enclosed in a metal box to eliminate noise coupled into the receiver from adjacent parts of the system into which they were designed. Systems that require metal shielding have proved to be unusual. Thus, in the majority of applications, the inherent noise immunity of the components combined with the shielding provided by the receiver ground plane have provided sufficient noise immunity.

150 MBd FIBER OPTIC RECEIVER FOR –5.2 V INTERFACE

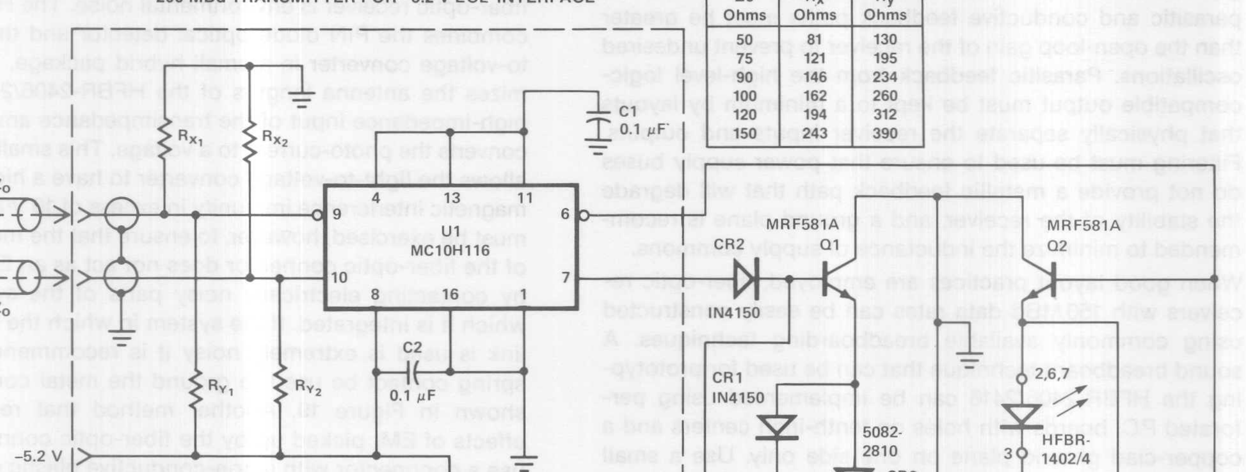


AMBIENT TEMPERATURE	25° C			
DATA FORMAT	2^{23}-1 PRBS			
DATA RATE (MBd)	150	130	100	50
RECEIVER SENSITIVITY (dbm_{PK})	26	27	29	30
DYNAMIC RANGE	18	19	21	22

NOTE:
V_{BB} IS A BIAS VOLTAGE GENERATED INTERNALLY BY THE 10116 ECL LINE RECEIVER.

Figure 16a.

200 MBd TRANSMITTER FOR –5.2 V ECL INTERFACE



Termination Resistor Values		
Z_O Ohms	R_X Ohms	R_Y Ohms
50	81	130
75	121	195
90	146	234
100	162	260
120	194	312
150	243	390

NOTES:

(1) THIS PORTION OF THE TRANSMITTER PROVIDES A CONVENIENT INTERFACE TO LABORATORY TEST EQUIPMENT AND IS NOT REQUIRED WHEN DIRECTLY INTERFACED TO ECL.

(2) FOR SINGLE ENDED OPERATION CONNECT UNUSED INPUT TO PIN 11 OF THE 10116.

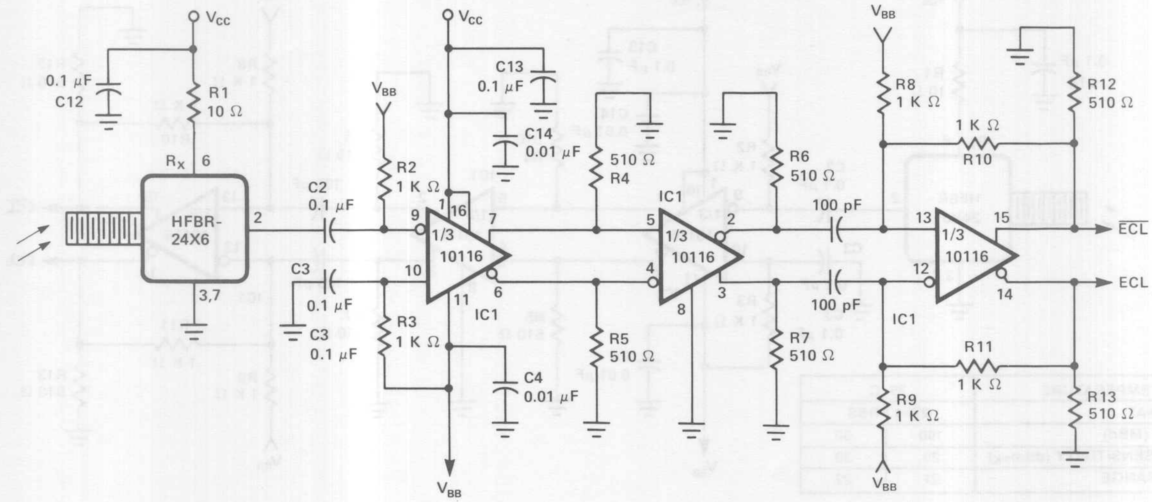
PEAK OUTPUT OPTICAL POWER MEASURED OUT OF 1 M OF CABLE

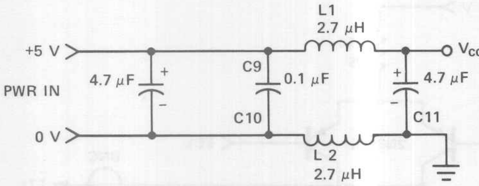
I_F = 60 mA T_A = 25°C

Fiber Cable	NA	HFBR-1402	HFBR-1404
HP 100/140 μm	0.3	–11.5 dBm	–7.1 dBm
62.5/125 μm	0.28	–16.5 dBm	–12.2 dBm
50/125 μm	0.20	–21.9 dBm	–17.5 dBm

Figure 16b.

150 MBd FIBER OPTIC RECEIVER FOR +5 V ECL INTERFACE TO AM 7969



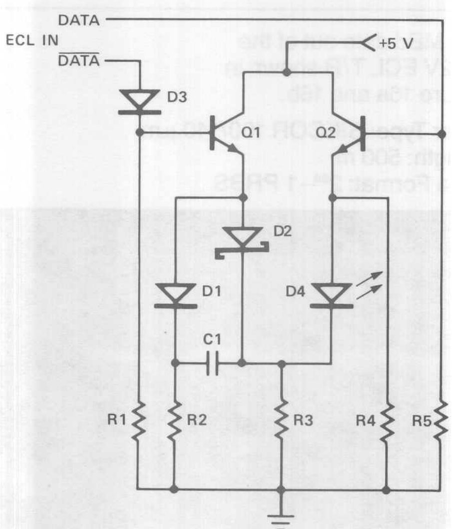


AMBIENT TEMPERATURE	25°C			
DATA FORMAT	2^{23}-1 PRBS			
DATA RATE (MBd)	150	130	100	50
RECEIVER SENSITIVITY (dbm_{PK})	26	27	29	30
DYNAMIC RANGE	18	19	21	22

NOTE: V_{BB} IS A BIAS VOLTAGE GENERATED INTERNALLY BY THE 10116 ECL LINE RECEIVER.

Figure 17a.

200 MBd TRANSMITTER CIRCUIT FOR +5V ECL
+5 V ECL INTERFACE TO AM 7968



Q1, Q2	MRF581A
D1, D3	IN4150
D2	5082-2810
D4	HFBR-14XX
C1	100 pF
R1, R5	215 Ω
R2	41.2 Ω
R3	31.6 Ω
R4	51.1 Ω

PEAK OUTPUT OPTICAL POWER MEASURED OUT OF 1 M OF CABLE

		I_F = 60 mA	T_A = 25°C
Fiber Cable	NA	HFBR-1402	HFBR-1404
HP 100/140 µm	0.3	-11.5 dBm	-7.1 dBm
62.5/125 µm	0.28	-16.5 dBm	-12.2 dBm
50/125 µm	0.20	-21.9 dBm	-17.5 dBm

Figure 17b.

(V) APPLICATIONS SUPPORT

Some complete designs that allow the use of HFBR-2406/2416 for run-length-limited data applications will now be discussed. Various interface requirements have been met which permit the HFBR-2406/2416 to be interfaced with:

1. ECL logic operating on -5.2 V. (Fig. 16)
2. The AMD TAXIchip™ +5 V 100 K ECL interface. (Fig. 17)
3. TTL logic operated on +5 V. (Fig. 18)

At an ambient temperature of 25°C all three interface circuits provided a typical receiver sensitivity of -32 dBm average with a BER of 1×10^{-9} at a data rate of 100 MBd. The TTL receiver is restricted to a maximum data rate of 100 MBd by the rise fall time of the ECL to TTL converter. All of the ECL interfaces are capable of operating at the 130 MBd data rate allowed by the guaranteed maximum rise time of the HFBR-2406/2416. Sensitivity at 130 MBd is typically -30 dBm average at a BER of 1×10^{-9}. Figure 20 shows the typical performance of the ECL transmitter/receiver at 25°C. Note that in this test a (2^{23})-1 PRBS pattern at 150 MBd was transmitted over 500 m of 100/140 µm graded-index fiber at a BER less than 1×10^{-9}.

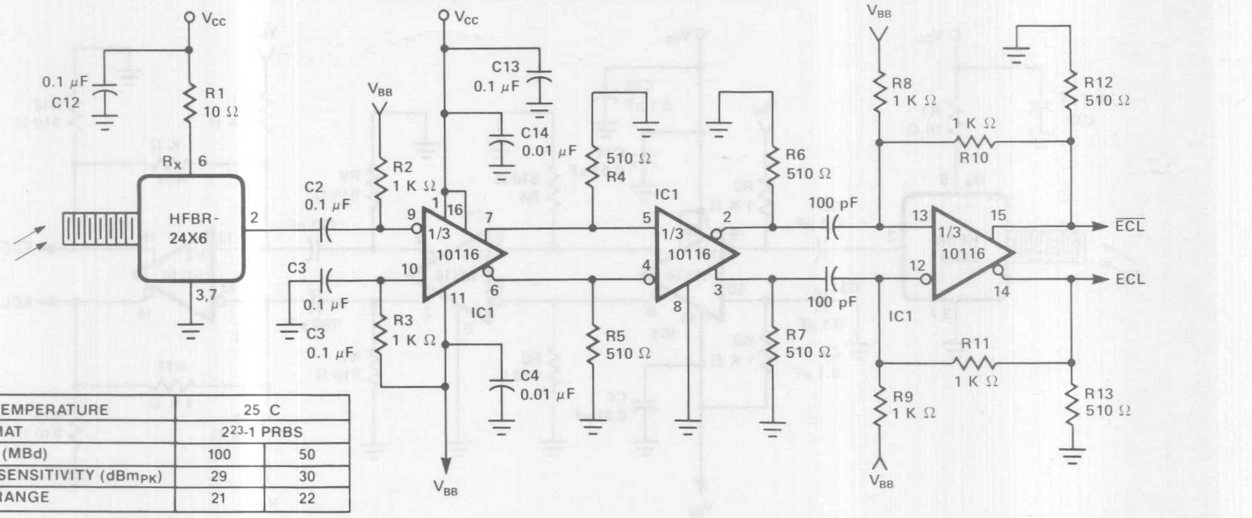
If the low-cost high-performance fiber-optic links possible with the HFBR-2406/2416 interest you, contact your local HP Field Sales Engineer for additional assistance. Your local HP sales representative can simplify your prototyping task by providing complete artwork for the fiber-optic transmitters and receivers discussed in this Applications Bulletin.

References

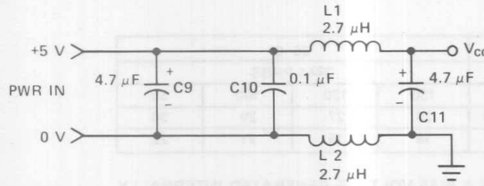
1. Hewlett-Packard Optoelectronics Designer's Catalog 1988, HFBR-ASWXXX data sheet.
2. James J. Refi "LED Bandwidth of Multimode Fibers as a Function of Laser Bandwidth and LED Spectral Characteristics", Journal of Lightwave Technology, Volume LT-4 No. 3 March 1986.
3. Delon C. Hanson and Jerry Hutchison "LED Source and Fiber Specification Issues for the FDDI Network", COMPCON Spring '87, IEEE Computer Society, (San Francisco, CA.), February 24-26, 1987.
4. Delon C. Hanson "Fiber Optic Sub System for Local Area Networks", OFC '88, (New Orleans, LA.), January 24-28, 1988.

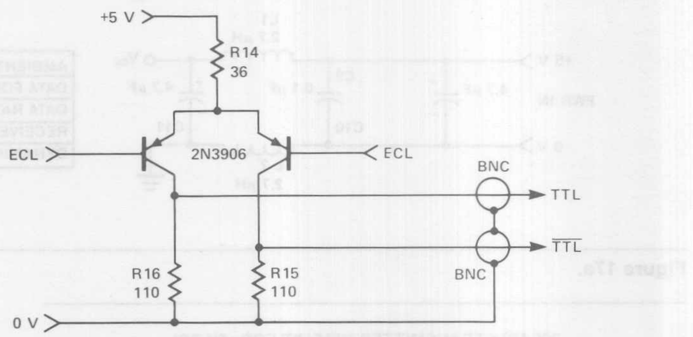
*TAXIchip is a registered trademark of Advanced Micro Devices, Inc.

100 MBd FIBER OPTIC RECEIVER FOR TTL INTERFACE



AMBIENT TEMPERATURE	25 C	
DATA FORMAT	2^{23}-1 PRBS	
DATA RATE (MBd)	100	50
RECEIVER SENSITIVITY (dBm_{PK})	29	30
DYNAMIC RANGE	21	22

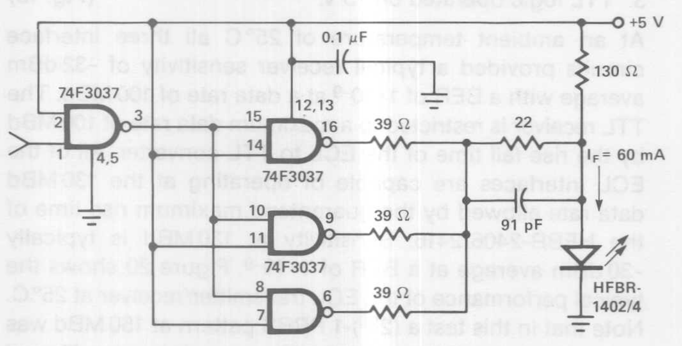




NOTES:
1. V_{BB} IS A BIAS VOLTAGE GENERATED IN THE 10116 INTEGRATED CIRCUIT THAT IS BROUGHT OUT AT PIN 11.
2. THIS RECEIVER IS INTENDED FOR USE WITH RUN LENGTH LIMITED DATA WHICH HAS A DUTY FACTOR BETWEEN 40 TO 60%.

Figure 18a.

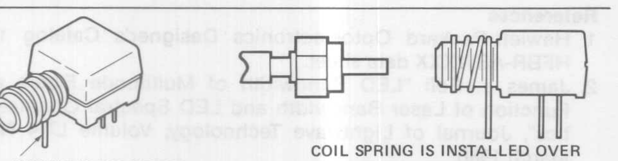
120 MBd TRANSMITTER FOR TTL INTERFACE



PEAK OUTPUT OPTICAL POWER MEASURED OUT OF 1 M OF CABLE

		I_F = 60 mA	T_A = 25°C
Fiber Cable	NA	HFBR-1402	HFBR-1404
HP 100/140 µm	0.3	-11.5 dBm	-7.1 dBm
62.5/125 µm	0.28	-16.5 dBm	-12.2 dBm
50/125 µm	0.20	-21.9 dBm	-17.5 dBm

Figure 18b.



SPRING IS SOLDERED TO RECEIVER GROUND PLANE.

COIL SPRING IS INSTALLED OVER THE OPTICAL PORT SO THAT F.O. CONNECTOR IS GROUNDED WHEN CONNECTED TO THE HFBR-2406/2416.

Figure 19.

150 MBd data out of the –5.2V ECL T/R shown in Figure 16a and 16b.

Fiber Type: SIECOR 100/140 µm
Length: 500 m
Data Format: 2^{23}–1 PRBS

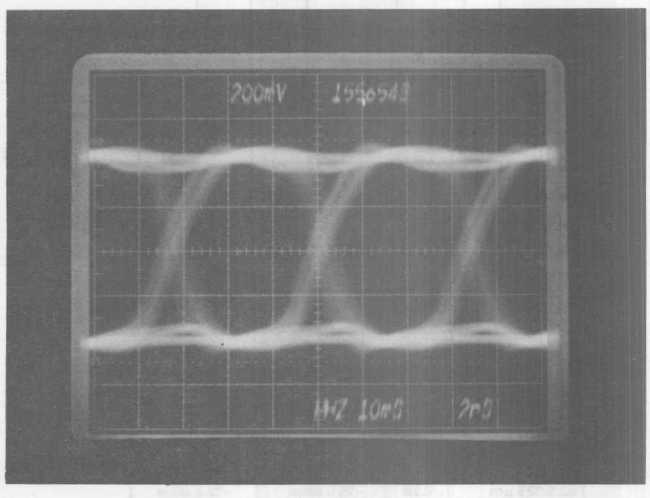


Figure 20.

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